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Theory of A Wideband Distributed Gyrotron Travelling Wave Amplifier

K. R. Chu, Y. Y. Lau, L. R. Barnett and V. L. Granatstein

Plasma Physics Division

June 23, 1981

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We present the concept of an ultra-wideband distributed gyrotron travelling wave amplifier for millimeter and submillimeter waves. The radius of the waveguide in the interaction region is increased along the axis while the strength of the d.c. magnetic field is decreased in such a way that the wave cutoff frequency is kept nearly equal to the electron cyclotron frequency. The basic principle of operation, peak gain, and saturated efficiency are analyzed. It is shown that instantaneous bandwidth over at least two octaves is theoretically possible. Technological requirements for achieving such an amplifier are assessed, including proposed structures for distributed input wave coupling.

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CONTENTS

THEORY OF A WIDEBAND DISTRIBUTED GYROTRON TRAVELLING WAVE AMPLIFIER

I. INTRODUCTION

Power levels achieved by gyrotron oscillators have constituted a revolution in power available from coherent electromagnetic sources at millimeter wavelengths,[1-7] and gyrotron oscillators have already been successfully applied in an energetic-effects application, the heating of controlled fusion research plasmas.[11,12] However information-carrying-systems such as radar and communications are better served by an amplifier with substantial instantaneous bandwidth rather than by an oscillator. A gyrotron travelling-wave-tube amplifier has been proposed[14,15] and initial theoretical analyses[24] and experimental tests[25-28] have demonstrated power levels an order of magnitude greater than available in conventional millimeter-waver travelling-wave-tube amplifiers together with an instantaneous bandwidth of several percent. A bandwidth on this order is of interest in a number of systems but still larger bandwidth would increase the usefulness of the gyrotron amplifier. Other authors have suggested gyrotron-like amplifiers which would achieve large bandwidth, viz. a gyrotwistron with tapered cavities and magnetic field[29] and a slow wave cyclotron amplifier[30-32] employing a non-relativistic bunching mechanism.[33,34] The presently proposed configuration is a modification of the gyrotron travelling wave amplifiers whose operation has already been successfully demonstrated[25-28] moreover, since it is a fast wave device, it should be less sensitive to the degrading effect of electron velocity spread.

A scheme in which the input signal is initially injected in the reverse direction of the tapered waveguide was proposed and analyzed by Lau[35], including the effect of velocity spread[36], for a proof-of-principle experiment. The experiment was expected to have approximately 15% small signal bandwidth and the observed bandwidth is 13%[37].

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An alternative side wall wave injection scheme potentially more suitable for high gain operation is discussed in the present paper.

II. MODEL AND ASSUMPTIONS

Figure 1a shows the side view of a circular cross section waveguide immersed in an applied magnetic field (B). An annular electron beam propagates to the right along the axisymmetric magnetic field lines. The electrons have a substantial part of its kinetic energy in the form of gyrational motion and, in contrast to the conventional travelling wave tubes, the *transverse* kinetic energy is to be converted into electromagnetic radiation through the cyclotron maser interaction. Figure 1b shows a cross sectional view of the waveguide and the electron beam. We assume (i) that the total length (L) of the waveguide is much longer than the interaction length (ΔL) defined as the length over which the cyclotron maser interaction takes place for a fixed input wave frequency; (ii) that the radius (r_w) of the waveguide and the amplitude of the applied magnetic field (Fig. 3c) vary slowly along the axis so that each interaction section is characterized by a distinct frequency [Eq. (4)]; and (iii) all the electrons have the same perpendicular velocity ($v_\perp$) and axial velocity (v_z) as they enter the waveguide, and their guiding centers are located on the circle of radius r_g.

The main element in the broadbanding scheme is that different portions of the waveguide amplifies different frequencies. As a TE_{mn} wave of certain frequency (ω) is launched from the left into the waveguide structure it will be amplified in a particular interaction section of the waveguide where its cutoff frequency closely matches the wave frequency.

All the gyro-TWA's reported so far have operated near the cutoff frequency of the waveguide. In the $\omega - k_z$ diagram, where k_z is the axial wavenumber, this implies that the waveguide characteristic curve

$$\omega^2 - k_z^2 c^2 - k_{mn}^2 c^2 = 0 \tag{1}$$

intersects the beam characteristic curve

$$\omega - k_z v_z - s\Omega_c = 0 \tag{2}$$

at or near a grazing point (Fig. 2), where $k_{mn} = x_{mn}/r_w$, x_{mn} is the n-th nonvanishing root of $J_m'(x) = 0$, $\Omega_c = eB/\gamma mc$, $\gamma = (1 - v_\perp^2/c^2 - v_z^2/c^2)^{-1/2}$, and s is the cyclotron harmonic number. The advantage for grazing intersection is two fold. First, it corresponds to a small wave number (k_z) and therefore mitigates the effect of beam velocity spread. Secondly, the beam curve [Eq. (2)] will not intersect the waveguide curve [Eq. (1)] on the negative k_z axis and therefore backward TE_{mn} mode will not be excited. Thus, in the present scheme, it is best to adjust the applied magnetic field profile such that grazing intersection is maintained throughout the waveguide. Equations (1) and (2) give the magnetic field (or the cyclotron frequency Ω_c) needed for grazing intersection,

$$\Omega_c = x_{mn}\, c/(s\gamma_z r_w) \tag{3}$$

and the wave frequency at the point of intersection

$$\omega = \gamma_z\, x_{mn}\, c/r_w, \tag{4}$$

where $\gamma_z \equiv (1 - v_z^2/c^2)^{-1/2}$. From Eq. (3), we obtain the condition for maintaining grazing intersection,

$$B\, r_w\, \gamma_z = B_0\, r_{w0}\, \gamma_{z0}, \tag{5}$$

where, here and also in subsequent equations, the subscript "0" denotes values at the entrance of the waveguide ($z = 0$).

III. CALCULATION OF PEAK GAIN AND SATURATION BANDWIDTH

On the basis of assumption (ii) above, we may approximate the peak gain (g_p) and saturated efficiency (η) in any interaction section by the analytical expressions derived in Ref. 19.

$$g_p \simeq \frac{19\pi}{\beta_z}\left[\frac{\nu J_{s-m}^2(r_g/r_w)\, J_s'^2(r_L/r_w)\, \beta_\perp^2}{\gamma\, \gamma_z^4 x_{mn}^2\, K_{mn}}\right]^{1/3} \frac{dB}{\text{free space wavelength}} \tag{6}$$

$$\eta \simeq \frac{1.6}{\gamma - 1}\left[\frac{\nu\gamma^2\gamma_z^2 J_{s-m}^2(r_g/r_w) J_s'^2(r_L/r_w)\, \beta_\perp^2}{x_{mn}^2 K_{mn}}\right]^{1/3} \tag{7}$$

where $K_{mn} = J_n^2(x_{mn})(1 - m^2/x_{mn}^2)$, $J_n(x)$ is the Bessel function of order n, $J_n'(x) = dJ_n(x)/dx$, $\beta_\perp = v_\perp/c$, $\beta_z = v_z/c$, r_L is the electron Larmor radius, and ν is a dimensionless electron beam density parameter defined as

$$\nu = N r_e , \tag{8}$$

where N is the number of electrons per unit length and $r_e = \mu_0 e^2/4\pi m = 2.8 \times 10^{-13}$ cm, is the classical electron radius.

In Eqs. (6) and (7), γ, K_{mn}, and x_{mn} are independent of the position in the waveguide, while the other parameter γ_z, ν, $\beta_\perp$, r_g/r_w, and r_L/r_w all vary along the waveguide. Thus, to evaluate g_p and η as functions of the axial position in the waveguide (or r_w), we need to express these beam parameters in terms of r_w and their initial values at $z = 0$. This can be readily done using Eq. (5) and the following conservation relations:

(1) conservation of electron magnetic moment,

$$\beta_\perp^2/B = \beta_{\perp 0}^2/B_0. \tag{9}$$

(2) conservation of electron flux,

$$\nu\beta_z = \nu_0\beta_{z0}. \tag{10}$$

(3) conservation of magnetic flux,

$$B\, r_g^2 = B_0\, r_{g0}^2. \tag{11}$$

We present the results directly.

$$\frac{B}{B_0} = \frac{\gamma_{z0}^2\beta_{\perp 0}^2 r_{w0}^2}{2\, r_w^2}\left[1 + \left(1 + \frac{4\, r_w^2}{\gamma^2\gamma_{z0}^2\beta_{\perp 0}^4 r_{w0}^2}\right)^{1/2}\right], \tag{12}$$

$$\gamma_z = \left(\frac{1}{\gamma^2} + \frac{\beta_{\perp 0}^2 B}{B_0}\right)^{-1/2}, \tag{13}$$

$$\frac{\beta_\perp}{\beta_{\perp 0}} = \left(\frac{B}{B_0}\right)^{1/2}, \tag{14}$$

$$\frac{r_L}{r_w} = \frac{r_{L0}}{r_w}\frac{B_0^{1/2}}{B^{1/2}}, \tag{15}$$

$$\frac{r_g}{r_w} = \frac{r_{g0}\, B_0^{1/2}}{r_w\, B^{1/2}}, \qquad (16)$$

$$\frac{v}{v_0} = \beta_{z0}\left(1 - \frac{1}{\gamma^2} - \frac{\beta_{\perp 0}^2 B}{B_0}\right)^{1/2}. \qquad (17)$$

Equations (12) through (17) allow us to express η, g_p, and B in terms of r_w, while r_w can in turn be expressed in terms of ω through Eqs. (5) and (12), and (13). Thus, we may express η, g_p, and B as a function of ω without specifying the explicit dependence of r_w on z. Figure 3 provides two examples showing the dependence of η, g_p, and B on ω over several octive bands for the TE_{01} and TE_{11} modes interacting with the beam at the fundamental of the electron cyclotron frequency. The electron beam energy is 70 keV with $v_{\perp 0}/v_{z0} = 1.5$ and all quantities in Fig. 3 except for η are normalized to their respective values at $z = 0$. The initial values of r_g are indicated in the figure caption. It is seen from Fig. 3a that the saturation bandwidth (defined as the interval of frequency between points with half of the peak efficiency) of more than two octaves are theoretically possible, especially for the TE_{01} mode. On the other hand, the peak gain decreases monotonically as the frequency decreases (Fig. 3b). This is expected because g_p is proportional to $\beta_z^{-1}\beta_\perp^{2/3}$, which decreases as the beam moves downstream along the decreasing magnetic field. Figure 3c shows that the applied magnetic field is approximately proportional to the wave frequency. Figure 4 provides examples for the second cyclotron harmonic interaction ($\omega \approx 2\Omega_c$), also exhibiting similar general characteristics as described above.

As shown in Eqs. (6) and (7), both the gain and efficiency are proportional to $J_{s-m}^{2/3}(r_g/r_w)$. Since Bessel function of zero order $[J_0(x)]$ has the largest amplitude, the highest gain and efficiency for the s-th cyclotron harmonic generally occurs for azimuthal waveguide modes with azimuthal mode number $m = s$. This characteristic of gyro-TWA is quantitatively exhibited in Figs 3 and 4.

We note here that Eqs. (6) and (7), obtained in Ref. 19, are accurate for all nonfundamental cyclotron interactions as well as the fundamental cyclotron interaction with beam energy $\geq$ 70 keV. But for the fundamental cyclotron harmonic interaction with beam energy below 70 keV, they lead to overestimates. The reason is that in the latter case, the free energy depletion saturation, which was neglected in Eqs. (6) and (7), becomes important.

Up to this point, r_w is still an unspecified function of z. We proceed to show how $r_w(z)$ can be determined under the requirement that the total gain (G) has a uniform value for all frequencies. We note that g_0 in Eq. (6) is the peak gain per unit length at the center of the interaction region. The actual gain (for a fixed frequency) tapers off on both sides. Assuming the interaction region for a fixed frequency extends from z_1 to z_2, the total gain is then given in terms of the local gain per unit length $g(z,\omega)$ by

$$G(\omega) = \int_{z_1}^{z_2} g(z,\omega)\, dz, \tag{18}$$

where z_1 and z_2 are determined by $g(z_1,\omega) = g(z_2,\omega) = 0$, and $g(z,\omega)$ may be evaluated from the dispersion relation [Eq. (8) of Ref. 19]. Thus, in principle, given a desired total gain, one may determine the waveguide profile, $r_w(z)$, from Eq. (18). As outlined above, the distributed nature of the amplification processes and the flexibility to shape the waveguide profile allow one to design an amplifier with uniform total small signal gain across the entire frequency band. The saturated bandwidth, however, has an intrinsic limit as shown in the preceding efficiency calculations.

It is worth noting that although a long waveguide is required for wideband operation, the interaction length for each frequency remains relatively short. Hence beam velocity spread, while reducing the gain and efficiency, does not pose any more difficulty in wideband operations that it would in narrow band operations. When this kind of broadbanding method is employed in gyrotwistrons[29] in which electron bunching and energy extraction take place in two separate sections, wider bandwidth necessitates greater separation between the two sections; velocity spread spoils the coherence as the separation increases and consequently limits the achievable bandwidth. In the optimized example of Ref. 29, for example, the calculated bandwidth is 7% for a beam with 10% velocity spread.

IV. A PROPOSED DISTRIBUTED INPUT COUPLER

In order to take full advantage of the very broadband nature of the distributed gyrotron amplifier, one will need to develop correspondingly broadband input couplers in a compatible geometry. One possiblity, the distributed input coupler, is described below.

The distributed input coupler is, in effect, a microwave multiplexer in which signal bands (or channels) are separated out of the common input line and injected at the appropriate position along the tapered interaction circuit. For this application, we propose a multiplexer-distributed coupler circuit that consists of a multiple of channel filters connected between the input rectangular waveguide operating in the fundamental TE_{10} mode and the taper interaction waveguide. The channel filters consist of coaxial cavities excited in a mode which will couple through apertures in the inner surface to excite the desired mode in the tapered waveguide. For efficient transmission through the cavity the input and output coupling is tight such that the loaded Q is much less than the unloaded Q.[38] The loaded Q will depend on the channel width desired (order of 1 to 2%). The circuit of Figure 5 consists of a multiple of single cavity filters connected by apertures to the input waveguide.[39,40] The cavities are tuned to separate center frequencies and are located on odd number of quarter wavelengths (i.e., $\lambda/4$, $3\lambda/4$, etc. of the respective cavity) from a short. In this case a resonant cavity acts as a shunt impedance to the input waveguide and the non-resonant cavities appear as open circuits and do not couple.[39] For use as an input coupler the bands f_1, f_2, etc. are injected into the tapered waveguide at the proper points for amplification.

With guard bands between the channels as normally would exist for this type of multiplexer, only one cavity is coupled at a time and the design is simple. However, since what is required is a multiplexer with contiguous pass bands (i.e., no guard bands) which typically cross over at the 3 dB points of the filters,[40] then two cavities will strongly couple near the cross over frequencies. In addition, the output of each cavity is recombined into the common tapered interaction waveguide circuit. The requirements of contiguous pass bands, recombination in the tapered circuit, low VSWR across the entire band, and good transmission efficiency will necessitate careful design, and may require additional matching elements, decoupling cavities, etc.

The cavities can be devised several ways. The suggested cavity for a TE_{21} amplifier is a TE_{411} coaxial cavity as shown in Figure 6a. Four azimuthal current maximums exist on the inner wall. Therefore, four axial slots apertures in the inner wall would couple strongly to TE_{21}, on the inside.

With this method, mode selectivity is good. As might be suspected, any of the lower modes can be excited by a coaxial cavity operating in the corresponding mode, i.e., a TE [illegible] will couple to a TE [illegible], TE_{0l} will couple to a TE [illegible], etc. The proper number and location of the axial slots must be employed, however, a TE [illegible] will couple to TE [illegible] and also to TE [illegible], if only 2 opposing coupling slots are used, for example. In the case of the fundamental mode, TE [illegible], single slot coupling may suffice and a simpler cavity (such as rectangular) could be used.

Resonant wavelengths of a full coaxial cavity are given by

$$\lambda = 2\left[\left(\frac{2\chi_{mn}}{\pi a}\right)^2 + \left(\frac{l}{L}\right)^2\right]^{-1/2}$$

where $m,n,$ and l correspond to the TE_{mnl} modes, L is the length, and a is the outer diameter. Plots of the root values χ_{mn} for a number of low order modes as a function of the ratio of the wall radii have been given[43] as well as formulas for the cavity Q's.

Although the lower order coaxial cavity modes are fairly wide-spaced, wide bandwidth amplifier designs will cross spurious resonances. Coupling apertures which minimize coupling to the spurious modes, loading of the spurious modes, fins, etc. are techniques which can be used to minimize spurious mode interference. We suggest that, instead of a single coaxial cavity being the filter element between the input and interaction circuit, several coupled cavities in tandem be used in which simple (such as rectangular) cavities precede and follow the coaxial cavity such as illustrated in Figure 6b. The added cavities would have spurious modes outside the amplifier band of interest and therefore isolate the coaxial cavity from the input and interaction waveguides. In addition, with appropriate coupling and stagger tuning of the cavities, filters can be made which have much better passband response, than the simple single cavity filter.

V. SUMMARY

A concept for modification of the gyrotron travelling-wave-amplifier which promises to result in extremely broadband operation has been presented. The modification consists of increasing the drift

tube radius as a function of axial position while at the same time decreasing the strength of the applied magnetic field so as to keep wave cutoff frequency nearly equal to the electron cyclotron frequency throughout the waveguide. The total linear gain can be made essentially independent of frequency by choosing optimized axial contours of wall radius and magnetic field. The saturated efficiency has been calculated for a number of modes with interaction at both the fundamental of the cyclotron frequency and at twice the cyclotron frequency, typically octave-like saturated bandwidths appear to be possible with a saturated bandwidth greater than two octaves calculated for both the TE_{01} and TE_{11} modes. A concept has also been presented for a distributed input coupler involving multi-cavity coupling between an input rectangular waveguide and the contoured drift tube; this type of input coupler promises to be compatible in bandwidth and in geometry with the broadband distributed gyrotron travelling-wave amplifier. A practical working model of the distributed input coupler remains to be developed.

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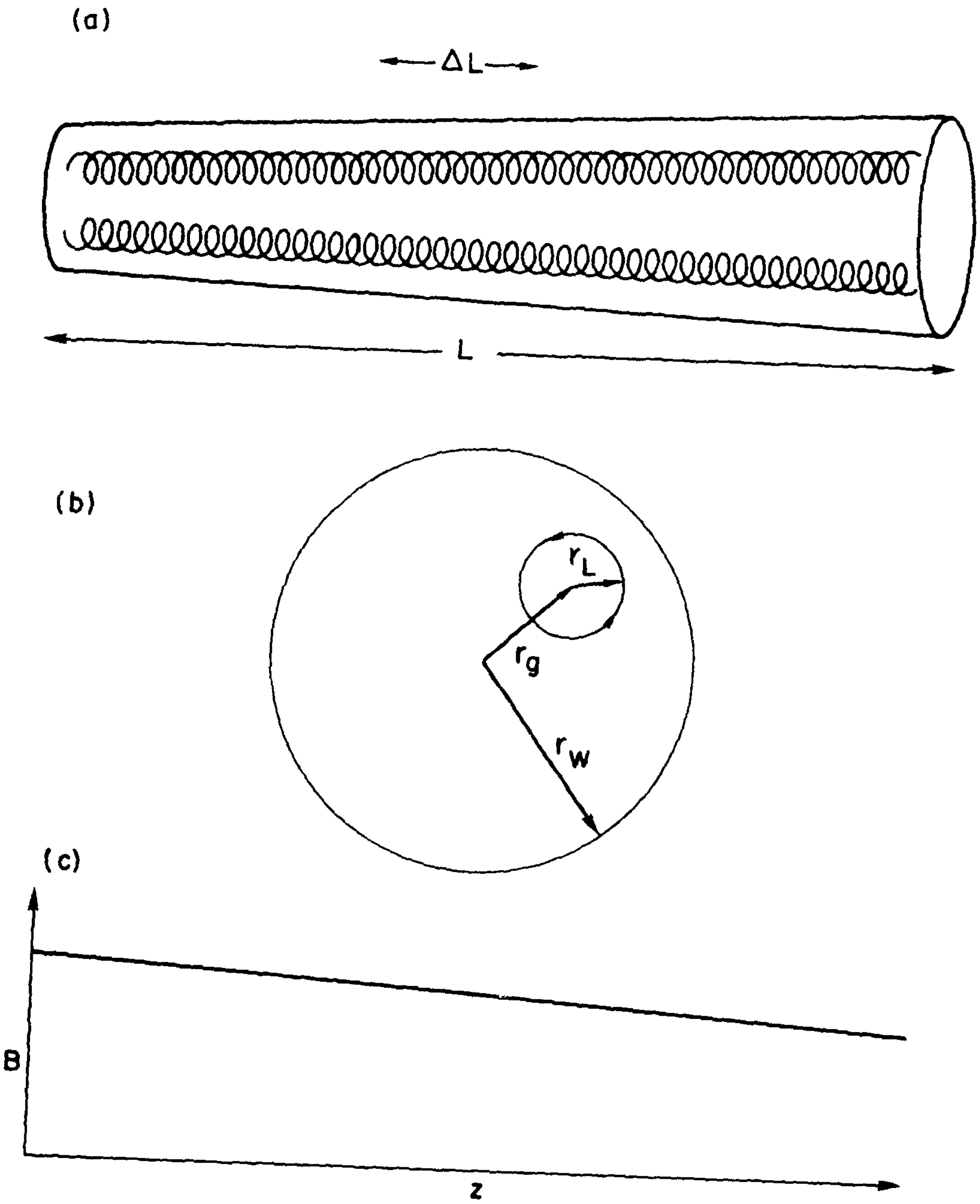

Fig. 1 — Side view (a), cross-sectional view (b), and the magnetic field profile (c) of the distributed gyrotron travelling wave amplifier. The main feature is the gradual tapering of the waveguide wall and the applied magnetic field. The degree of tapering is exaggerated in the figure. The actual interaction takes place over a distance ΔL much smaller compared with the total length (L) of the system. Location of the interaction region varies with the wave frequency.

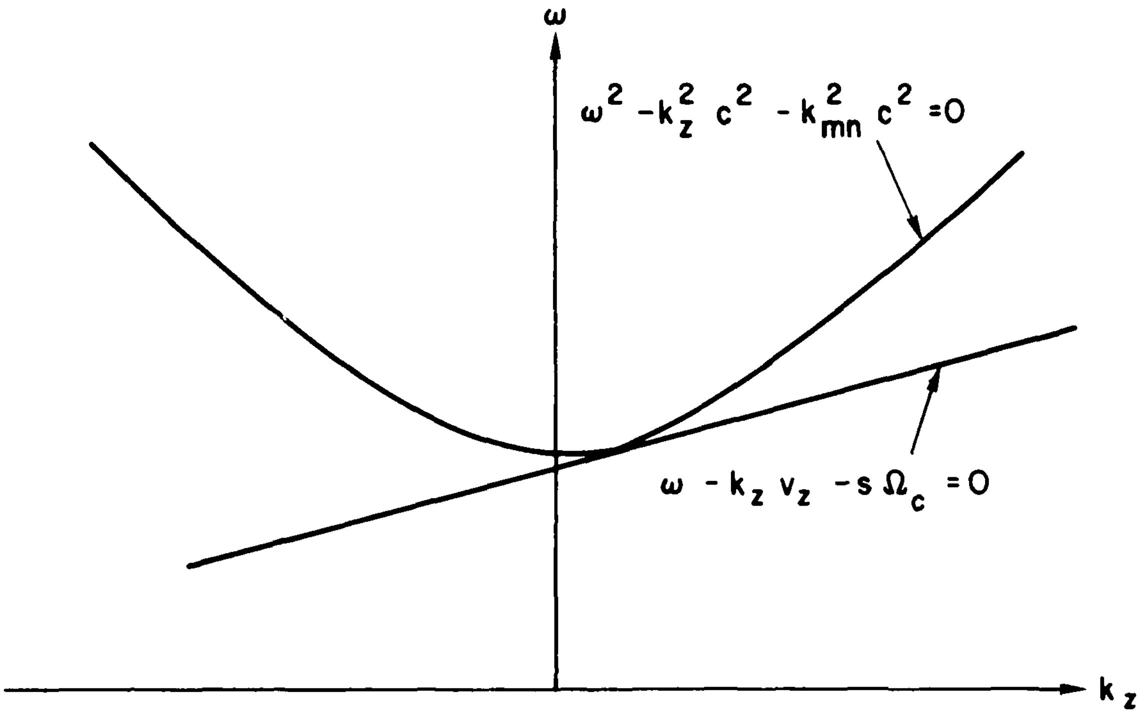

Fig. 2 — The waveguide wall and the applied magnetic field are tapered in such a way that at any section of the waveguide, the waveguide mode [Eq. (1)] aways intersects with the beam mode [Eq. (2)] at or near the grazing point

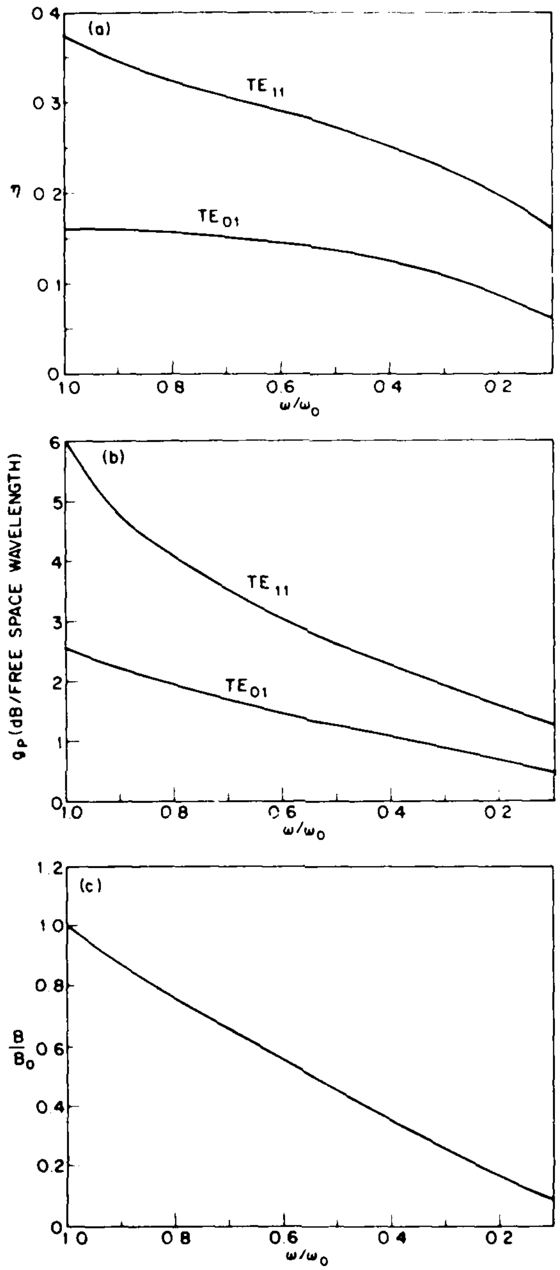

Fig. 3 — Plots of efficiency (η), peak gain (g_p) and applied magnetic field (B) versus ω for the TE_{11} modes and the first cyclotron harmonic interaction. The beam parameters are $\gamma = 1.14$ (70 keV) and $v_{\perp 0}/v_{\parallel 0} = 1.5$. The initial beam guiding center positions and Larmor radii are $r_{c0}/r_{w0} = 0.7$, $r_{L0}/r_{w0} = 0.11$ for the TE_{01} mode, and $r_{c0}/r_{w0} = 0.3$, $r_{L0}/r_{w0} = 0.23$ for the TE_{11} mode. B and ω are normalized to their values at $z = 0$.

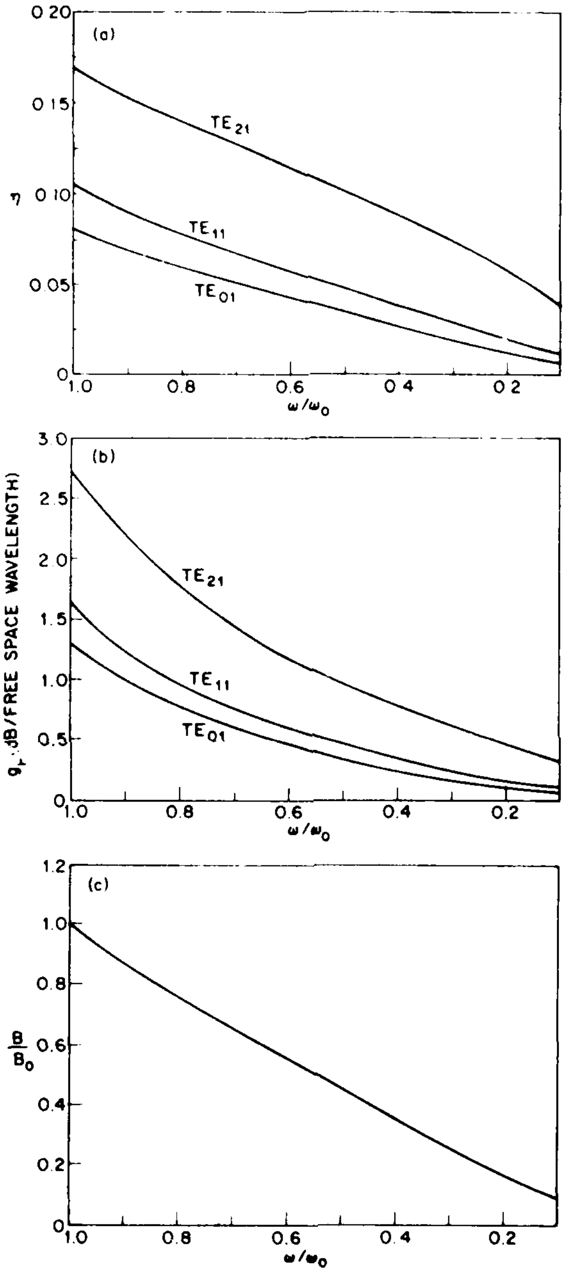

Fig. 4 — Plots of η, g_p, and B versus ω for the TE_{01}, TE_{11} and TE_{21} modes and the second cyclotron harmonic interaction. Beam parameters are the same as in Fig. 3. The initial beam guiding center positions and Larmor radii are $r_{g0}/r_{w0} = 0.6$, $r_{L0}/r_{w0} = 0.22$ for the for the TE_{01} mode, $r_{g0}/r_{w0} = 0.4$, $r_{L0}/r_{w0} = 0.48$ for the TE_{11} mode, and $r_{g0}/r_{w0} = 0.2$, $r_{L0}/r_{w0} = 0.27$ for the TE_{21} mode.

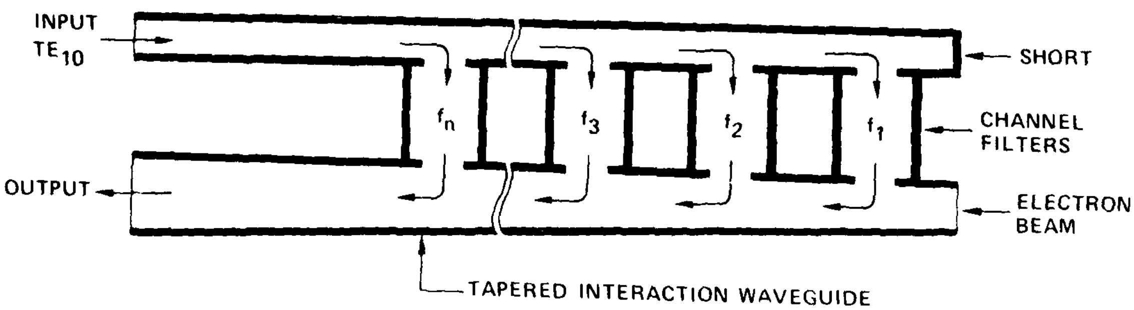

Fig. 5 — Concept of the proposed distributed input coupler with simple single cavity channel filters

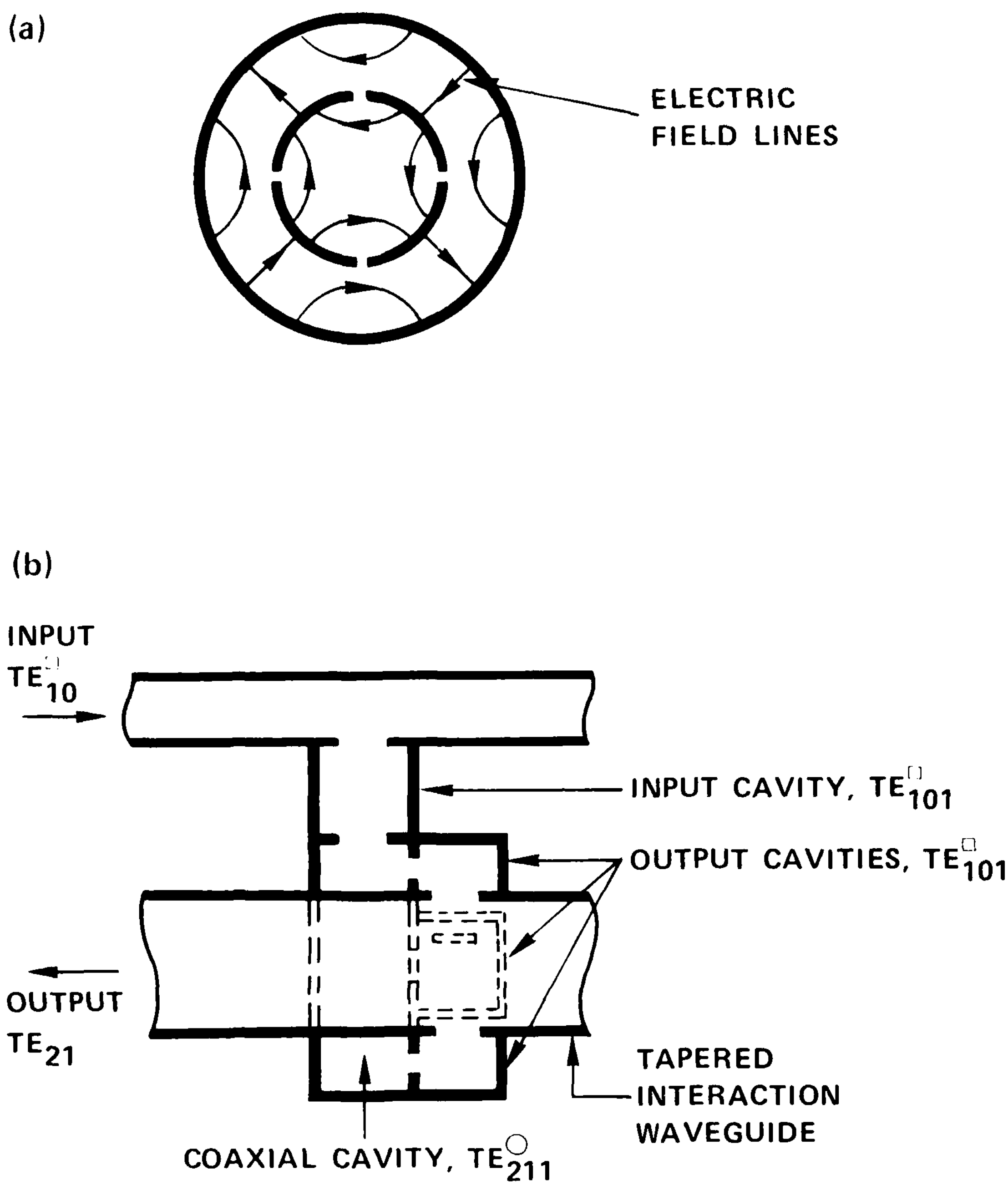

Fig. 6 — (a) TE_{211} coaxial mode with coupling apertures to the inner circular waveguide to excite TE_{21} mode (b) the form of the proposed channel filter for a TE_{21} mode amplifier

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